Novel Approach to Design Ultra Wideband Microwave Amplifiers: Normalized Gain Function Method

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Abstract. *In this work, we propose a novel approach called "Normalized Gain Function (NGF) method" to design low/medium power single stage ultra wide band microwave amplifiers based on linear S parameters of the active device. Normalized Gain Function T_{NGF} is defined as the ratio of T and $|S_{21}|^2$, desired shape or frequency response of the gain function of the amplifier to be designed and the shape of the transistor forward gain function, respectively. Synthesis of input/output matching networks (IMN/OMN) of the amplifier requires mathematically generated target gain functions to be tracked in two different nonlinear optimization processes. In this manner, NGF not only facilitates a mathematical base to share the amplifier gain function into such two distinct target gain functions, but also allows their precise computation in terms of $T_{NGF}=T/|S_{21}|^2$ at the very beginning of the design. The particular amplifier presented as the design example operates over 800-5200 MHz to target GSM, UMTS, Wi-Fi and WiMAX applications. An SRFT (Simplified Real Frequency Technique) based design example supported by simulations in MWO (MicroWave Office from AWR Corporation) is given using a 1400 mW pHEMT transistor, TGF2021-01 from TriQuint Semiconductor.*

Keywords

Ultra wideband microwave amplifiers, Simplified Real Frequency Technique (SRFT), small signal scattering parameters, matching networks, Chebyshev approximation, nonlinear optimization, Matlab, Microwave Office (AWR).

1. Introduction

Mobile wireless equipment of today have unnecessarily large number of narrow-band power amplifiers (PA) in their output stages to accommodate communication standards such as GSM, DCS1800, PCS1900, UMTS, Bluetooth, WLAN, Wi-Fi and WiMAX. The number of PAs reaches even up to 6 especially in some GSM platforms. In this work, based on a very recently developed novel ultra wide band (UWB) amplifier design approach that we call "*NGF (Normalized Gain Function) method*" [1], we propose to replace a large number of PAs by designing only one single UWB PA operating within 800-5200 MHz band. Therefore, the use of this type UWB amplifier decreases the problems such as system complexities, high costs, heavy equipment structure, large circuitry areas, high DC power consumption caused by many separate narrow-band PAs and their accompanying matching elements. NGF based design methodology is used to design UWB amplifiers operating within 800-5200 MHz band that can cover all frequency bands of the above mentioned communication standards.

Increasing sophistication in modern wireless equipment tends to require more and more digital hardware and software units to control output amplifier of the transmitter especially in SDR (Software Defined Radio) based platforms. From the flexible operation point of view, a highly digital environment can benefit having such a UWB amplifier. In such a SDR platform, signals to be broadcasted are generated via only software means in the digital unit (FPGA, DSP etc.) and directly applied into the input port of the proposed UWB amplifier without being exposed any drawbacks by hardware intervention. However, from the embedded engineering point of view, digitally generated signals must be safely targeted to fall into the operational frequency band of the UWB amplifier.

Due to impossibility of analytically solving the amplifier equations composed of many unknown variables related to element values of the input and output matching networks (IMN/OMN), one has to choose a numerical technique that is highly successful, numerically stable and yielding always convergent and realizable solution. The obvious answer is the Real Frequency Techniques (RFTs) which are very well-known and widely used numerical solvers in the literature [2-8]. In this work, reflectance based RFT version called "Simplified Real Frequency Technique (SRFT)" is preferred as the fundamental numerical design technique [4-20].

NGF takes its name from the amplifier gain function T divided by the forward scattering parameter of the transistor $|S_{21}|^2$, a kind of normalization or division operation applied to the amplifier gain function T by the normalization factor $|S_{21}|^2$. We may also use a term NGF-SRFT (NGF assisted SRFT technique) suitably which combines NGF method with SRFT based design technique.

An UWB Flat Gain Amplifier (FGA) with a flat gain level $T_0 = 14.90$ dB (see Fig. 1.b) along $[f_L - f_H]$ = [0.8-5.2] GHz frequency band is designed using the developed Matlab [21] code. The linear small signal normalized scattering parameters of a commercially available 1400mW wideband DC-12 GHz discrete power pHEMT, TGF2021-01 of TriQuint Semiconductor [22], is used in the design and the theoretical results are found in 100% agreement with the simulations done in the Microwave Office (MWO) of AWR Corporation [23]. TGF2021-01 is ideally suited for point-to-point radio, high-reliability space, and military applications [22], however in such applications demanding more than 1400 mW power, the introduced design method allows us to design more powerful amplifiers using higher power transistors provided that they are targeted to operate in the linear region sufficiently below P_{1dB} point.

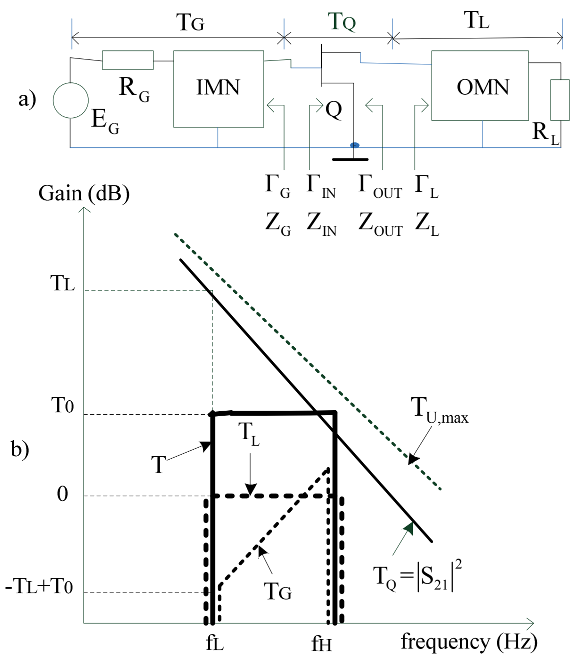


Fig. 1. a) An ultra wideband microwave amplifier composed of IMN, OMN, a single transistor Q, source and load side resistive terminations RG and RL. b) Gain curve shapes for the amplifier operating in the (f_L-f_H) frequency band: T_G, $T_Q = |S_{21}|^2$, T_L, T, $T_{U,max}$ (Maximum Unilateral Power Gain, i.e. $\Gamma_L = S_{22}^*$, $\Gamma_G = S_{11}^*$, if $S_{12} = 0$ assumed).

Major features of the NGF-SRFT methodology: NGF method to be presented in this paper has several novelties when compared to RFT based amplifier design methods worked in for example [5] and [10]:

i) NGF method starts to design OMN first and then IMN as opposed to them.

ii) It predetermines the shape of each target gain functions for OMN and IMN along the optimization frequency band and generates target gain functions modeled by bandpass Chebyshev or Butterworth type template functions. Thus, the optimization has gain curves to track precisely in the whole frequency band covering also stopband beside the passband. However, previous art uses target gain curves limited only in the passband.

iii) Bandpass Chebyshev or Butterworth modeled target gain curves behave as precise guides to be tracked by the optimization. Therefore, the search space is predetermined which makes the optimization is always convergent, numerically robust, well-behaved and capable to reach realizable solutions.

iv) NGF is general, i.e. it assumes no unilateral behavior known as $S_{12} = 0$ for the transistor. It takes into account of the output into the input or vice versa, by assuming $S_{12} \neq 0$. This causes transistor input and output reflectances be affected from each other which is not so in the case $S_{12} = 0$ assumption of previous works. This allows more realistic resemblance of the transistor in the design.

v) NGF takes into account of the acceptability of reflectance values when designing OMN and IMN. Reflectances should have sufficiently low values and this is tried to be achieved via LRA (Least Reflection Approach) or BRA (Balanced Reflection Approach) approaches which will be detailed in Section 3.1.B.

In the following sections, first, amplifier design equations are introduced in Section 2. In Section 3, Normalized Gain Function method is introduced. A brief summary of SRFT is outlined in Section 4. In Section 5, wideband amplifier design steps using NGF-SRFT are given and elaborated. Section 6 deals with an example design of an UWB amplifier based on NGF-SRFT and its performance evaluations.

2. Amplifier Design Equations

For the wideband microwave amplifier seen in Fig. 1.a which is composed of Input Matching Network (IMN), Output Matching Network (OMN), transistor Q, resistive terminations R_G and R_L at the generator (source) and load sides respectively, Transducer Power Gain (TPG or shortly T) is given as [24],

$$T = T_G T_Q T_L \tag{1}$$

where the partial gains are expressed in terms of IMN input reflectance Γ_G, reflectances Γ_{IN}, Γ_{OUT} seen from the input and output of the transistor and OMN input reflectance Γ_L such that [24],

$$T_G = \frac{1-|\Gamma_G|^2}{|1-\Gamma_{IN}\Gamma_G|^2},\ T_Q = |S_{21}|^2,\ T_L = \frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2}. \tag{2}$$

The reflectances seen from the input and output of the transistor can be written in terms of unit normalized scattering parameters (shortly S-parameters) given for the transistor, input reflectances Γ_G and Γ_L of IMN and OMN matching circuits as [24],

$$\Gamma_{IN} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L},\ \Gamma_{OUT} = S_{22} + \frac{S_{12}S_{21}\Gamma_G}{1-S_{11}\Gamma_G}. \quad (3)$$

Unit normalized scattering parameters are measured via a NA (Network Analyzer) under standard 50 Ohm resistive terminations connected at both ports of the transistor such that $R_1 = R_2 = 50$ Ohms, which corresponds to unit terminations $r_k = R_k/R_0 = 1$ Ohm ($k = \{1,2\}$) for a port normalization number $R_0 = 50$ at each ports. Both Γ_G and Γ_L are frequency dependent complex quantities expressed as rational functions defined by the ratio of two polynomials such that

$$\Gamma_G = \frac{h_G(p)}{g_G(p)},\ \Gamma_L = \frac{h_L(p)}{g_L(p)} \quad (4)$$

where $p = j\omega$ is the Laplace variable defined in terms of real frequency ω. Denominator polynomials for both reflectances must be "*Strictly Hurwitz"* polynomials whose all roots reside in the open LHP (Left Half Plane) and both numerator polynomials are arbitrary polynomials with real coefficients. Numerator and denominator polynomials of both matching networks are expressed in the following forms

$$\begin{aligned} &h_M(p) = h_{1,M}p^n + h_{2,M}p^{n-1} + \ldots + h_{n,M}p + h_{n+1,M} \\ &g_M(p) = g_{1,M}p^n + g_{2,M}p^{n-1} + \ldots + g_{n,M}p + g_{n+1,M} \\ &for\ M = \{G\ or\ L\} \end{aligned} \quad (5)$$

where M denotes generator (G) or load (L) side matching networks, i.e. IMN or OMN, respectively. n is the total number of L and C (inductors and capacitors) lumped elements in each matching network which has (n + 1) number of total elements including one resistive termination [7], [8].

Once either reflectance function is obtained in the form of (4), its corresponding Darlington driving point input impedance function is obtained as

$$\begin{aligned} z_M(p) &= \frac{1+\Gamma_M(p)}{1-\Gamma_M(p)} = \frac{g_M(p)+h_M(p)}{g_M(p)-h_M(p)} = \frac{a_M(p)}{b_M(p)} \\ &= \frac{a_{1,M}p^n + a_{2,M}p^{n-1} + \ldots + a_{n,M}p + a_{n+1,M}}{b_{1,M}p^n + b_{2,M}p^{n-1} + \ldots + b_{n,M}p + b_{n+1,M}},\ for\ M = \{G\ or\ L\} \end{aligned} \quad (6)$$

which must be a rational PRF (Positive Real Function) from the realizability point of view [7], [8].

The sole meaning of amplifier design is basically to determine the topology and circuit parameter values of IMN and OMN matching circuits, that is to realize [7], [25], [26] them using LC lossless components either in lumped, distributed or mixed fashion so that the resulting overall amplifier gain function T satisfy the prescribed gain curve shape such as seen in Fig. 1.b. Any PRF, thus "realizable", input impedance function $z(p)$ in the form of (6) can be synthesized in such a way that it always yields an LC network with resistive termination, which completes the design. This is known as "Darlington Synthesis" and it could be achieved via a procedure known as *Long-Division* or *Continued Fractional Expansion* using the following form [7], [8]

$$z(p) = z_1 + \cfrac{1}{+y_2 + \cfrac{1}{z_3 + \cfrac{1}{y_4 + \cfrac{1}{\ldots + \cfrac{1}{z_n(or\ y_n) + \xi_{n+1}}}}}} \quad (7)$$

where z_i and y_i designate the series arm impedances and the shunt arm admittances in a reciprocal lossless two-port ladder topology, respectively. ξ is the constant resistive termination (resistance or conductance) of the lossless two-port which can be omitted with a transformer loaded by unit resistance and

$$\{z_i\ and\ y_i\} = \begin{cases} a_i p \\ \dfrac{1}{b_i p} \\ \dfrac{c_i p}{c_i d_i p^2 + 1} \end{cases} \quad (8)$$

where coefficients a_i, b_i and "c_i-d_i pair" could be realized by inductor in series branch (capacitor in shunt branch), capacitor in series branch (inductor in shunt branch), "capacitor-shunt-inductor in series branch (capacitor-series-inductor in shunt branch)" if we deal with impedance z_i (or admittance y_i) function. The term related "c_i-d_i pair" represents a resonant circuit composed of serial or parallel connected L and C components [7], [8].

Losslessness Condition of Two-Port Networks: Any lossless two-port network must satisfy the relation known as "*losslessness condition*" given by

$$\begin{aligned} &g(p)g(-p) = h(p)h(-p) + f(p)f(-p) \\ &G(p) = H(p) + F(p) \end{aligned} \quad (9), (10)$$

which requires an extra polynomial $f(p)$ representing the transmission zeros of the gain function $T(p)$ of the two-port such that

$$f(p) = f_1 p^n + f_2 p^{n-1} + \ldots + f_n p + f_{n+1}. \quad (11)$$

$h(p)$ and $g(p)$ are numerator and denominator polynomials of the input reflectance function $\Gamma(p) = h(p)/g(p)$ and they have open form as given previously in (5). G, H, F of (10) are even polynomials each constructed by the product of corresponding polynomial and its conjugate. Conjugate polynomial is the polynomial in which Laplace variable is substituted by $-p$. These even polynomials are given with the following form

$$G(p)=G_1p^{2n}+G_2p^{2(n-1)}+...+G_np^2+G_{n+1}\ , \quad (12)$$

$$H(p)=H_1p^{2n}+H_2p^{2(n-1)}+...+H_np^2+H_{n+1}\ , \quad (13)$$

$$F(p)=F_1p^{2n}+F_2p^{2(n-1)}+...+F_np^2+F_{n+1} \quad (14)$$

where their order are double of the total element number n of the related matching network [4]. Gain function $T(p)$ of a lossless two-port (IMN or OMN) can be written in terms of scattering parameters of this two-port as $T_M(p)=\left|M_{21}(p)\right|^2=1-\left|M_{22}(p)\right|^2$, $M=\{G\ or\ L\}$, where $M_{21}(p)$ is the forward scattering parameter and $M_{22}(p)$ is the reflectance seen at port 2. As seen in Fig. 1.a, we use $\Gamma_M(p)$ notation instead of $M_{22}(p)$, hence the gain for either IMN or OMN can be written as $T_M(p)=1-\left|\Gamma_M(p)\right|^2$, $M=\{G \text{ or } L\}$. Therefore, for any lossless two-port, whether it is IMN or OMN, gain function is expressed in terms of even polynomials as

$$T(p)=1-\left|\Gamma(p)\right|^2=1-\frac{H(p)}{G(p)}=\frac{F(p)}{G(p)}\ , \quad (15)$$

$$T(p)=\frac{F(p)}{G(p)}=\frac{F_1p^{2n}+F_2p^{2(n-1)}+...+F_np^2+F_{n+1}}{G_1p^{2n}+G_2p^{2(n-1)}+...+G_np^2+G_{n+1}} \quad (16)$$

which is an even rational function [7], [8]. In many cases, realization of transmission zeros of the numerator polynomial given in general form (16) is difficult and mostly not practical. Therefore, one should adhere to a more practical model of the gain function given by

$$T(p)=\frac{F(p)}{G(p)}=\frac{F_0p^{2ndc}\prod_{i=1}^{n_z}\left(p^2+\omega_{z_i}{}^2\right)^2}{G_1p^{2n}+G_2p^{2(n-1)}+...+G_np^2+G_{n+1}} \quad (17)$$

which has ($2ndc + 2n_z$) number of total finite transmission zeros at DC and at frequencies ω_{Z_i}, $2n$-($2ndc$+$2n_z$) number of transmission zeros at infinity. In many practical cases, and for our work in this study which considers ladder topology only, we deal with a simplified gain function form as

$$T(p)=\frac{F(p)}{G(p)}=\frac{F_0p^{2ndc}}{G_1p^{2n}+G_2p^{2(n-1)}+...+G_np^2+G_{n+1}} \quad (18)$$

which only realizes transmission zeros at DC and infinity. Thus, this gain function form yields only single inductors and single capacitors in series/shunt branches of the ladder. In the paper, we utilize (18) to realize bandpass Chebyshev and Butterworth type gain functions. However, even though it is out of the scope of this study, one must also deal with finite real frequency or $j\omega$ zeros of $\prod_{i=1}^{n_z}\left(p^2+\omega_{Z_i}{}^2\right)^2$ term seen in (17) that may be realized either as a parallel resonance circuit in series configuration, or as a series resonance circuit in shunt configuration or even as a Darlington C (or D) section with coupled coils [7]. Especially, realization of Darlington C (or D) section with coupled coils is indeed a challenging issue and found impractical by the designers.

3. Normalized Gain Function Method

Shape or frequency response of amplifier gain function T in (1) is mathematically generated at the very beginning of the design in such a way that it exactly represents the desired technical specifications of the amplifier to be designed. Hence, it has a known shape $T_{data}(\omega)$ along the design or operating band $[f_b...f_e]$ of the amplifier. Moreover, the shape of the transistor forward gain function is also known as $T_{Q-data}(\omega)=\left|S_{21}(\omega)\right|^2$ since the transistor S-parameters are obtained from the vendor. Since shapes of these two functions are known, their ratio is a function whose shape is known as well. We define this ratio as "*Normalized Gain Function (NGF)*" [1] and write it as

$$T_{NGF-data}(\omega)=\frac{T_{data}(\omega)}{\left|S_{21}(\omega)\right|^2}=T_{G-data}(\omega)T_{L-data}(\omega)\ . \quad (19.a)$$

For a flat gain amplifier whose gain $T_{data}(\omega)=T_0$ as seen in Fig. 1.b, (19.a) becomes

$$T_{NGF-data}(\omega)=\frac{T_0}{\left|S_{21}(\omega)\right|^2}\quad ,\omega_b\le\omega\le\omega_e \quad (19.b)$$

which has a "tapered" gain shape due to reciprocal of $\left|S_{21}(\omega)\right|^2$ function [1]. Major objective of the designer is to able to obtain input and output matching networks (IMN/OMN) in such a way that the resulting amplifier satisfies the desired overall $T_{data}(\omega)$ gain shape seen in Fig. 1.b, provided that the maximum power transfer is attained as much as possible.

In the course of the amplifier design, designer targets to obtain matching networks IMN and OMN by means of two distinct optimization phases: *I. OMN design phase, II. IMN design phase,* respectively.

I. Output Matching Network (OMN) Design Phase:

OMN design phase aims to find OMN input reflectance $\Gamma_L(p)=h_L(p)/g_L(p)$ as a rational function in terms of h_L and g_L polynomial coefficients via a nonlinear optimization process. At the beginning of the design, the shape of the target gain function $T_{L-data}(\omega)$ to be tracked by the optimization algorithm is numerically generated. Then, unknown h_L and g_L polynomial coefficients of the input reflectance is obtained in the OMN optimization code via a suitable nonlinear optimization algorithm by minimizing the error or distance function given as

$$d_{OMN}(i)=T_L(p_i)-T_{L-data}(\omega_i),\quad i=1,2,...,nd \quad (20)$$

where data number in the frequency axis is *nd* and the minimum angular frequency step $d\omega = (\omega_e - \omega_b)/nd$. $p_i = j\omega_i$ is the numeric value of the Laplace variable at the i^{th} frequency point in the optimization or design band $[\omega_e ... \omega_b]$. In (20), OMN gain function in terms of unknown reflectance is evaluated at the i^{th} frequency point using

$$T_L(p_i) = \frac{1 - |\Gamma_L(p_i)|^2}{|1 - S_{22}(\omega)\Gamma_L(p_i)|^2}, \ \omega_b \le \omega_i \le \omega_e \ . \quad (21)$$

At the end of a successful optimization which may employ least square sense error minimization as $\sum_{i=1}^{nd} d_{OMN}(i)^2$, $z_L(p) = a_L(p)/b_L(p)$ is computed from the optimized $\Gamma_L(p) = h_L(p)/g_L(p)$ and synthesized in the form of (7) to get the topology and the element values of the OMN network, which completes the 1st phase of the design [1].

II. Input Matching Network (IMN) Design Phase:

Similar to the OMN design phase, the optimization code minimizes the IMN error function given as

$$d_{IMN}(i) = T_G(p_i) - T_{G-data}(\omega_i), \quad i = 1,2,...,nd \ . \quad (22)$$

IMN gain function in terms of unknown input reflectance is written as

$$T_G(p_i) = \frac{1 - |\Gamma_G(p_i)|^2}{|1 - \Gamma_{IN}(\omega_i)\Gamma_G(p_i)|^2}, \ \omega_b \le \omega_i \le \omega_e \quad (23.a)$$

where

$$\Gamma_{IN}(\omega_i) = S_{11}(\omega_i) + \frac{S_{12}(\omega_i)S_{21}(\omega_i)\Gamma_L(p_i)}{1 - S_{22}(\omega_i)\Gamma_L(p_i)} \ . \quad (23.b)$$

Similarly, at the end of a successful optimization which may employ least square error minimization of $\sum_{i=1}^{nd} d_{IMN}(i)^2$, $z_G(p) = a_G(p)/b_G(p)$ is computed from the optimized $\Gamma_G(p) = h_G(p)/g_G(p)$ and synthesized in the form of (7) to get the topology and element values of the IMN network, which completes the 2nd phase and the overall design [1].

3.1 Generation of Target Gain Functions

From the optimization point of view, error functions in (20) and (22) require pre-determination of shapes of $T_{L-data}(\omega)$ and $T_{G-data}(\omega)$ target gain functions before the optimization. The shapes of these target gain functions can be generated using two approaches: *A. Least Reflection Approach (LRA), B. Balanced Reflection Approach (BRA).*

A. Least Reflection Approach (LRA): In LRA [1], we propose to choose the OMN target gain function to satisfy the condition

$$|T_{L-data}(\omega)| = \alpha \le \alpha_{\max}, \ \omega_b \le \omega_i \le \omega_e \quad (24)$$

which makes OMN circuit have the “least reflection” or “reflectionless” (corresponds to $\Gamma_L = S_{22}^*$ along the passband in the ideal case or a very small value sufficiently close to zero in a well-matched network when $\alpha = \alpha_{\max}$) that preserve the transistor output from harmful load reflections especially in high power levels. Indeed, this is a critical task that must be charged on the transistor output circuit together with the OMN circuit through to the load since the transistor drain (collector) circuit has the capability of carrying high powers compared to the gate (base) circuit. A good candidate for OMN target gain function can be a bandpass gain function modeled by Chebyshev or Butterworth template function denoted by $T_{Ch}(\omega)$ or $T_{Bw}(\omega)$ of degree 2*n* and ripple factor ε with a maximum gain amplitude of unity as in (24) within $\omega_b \le \omega \le \omega_e$ band [1]. Therefore, using bandpass Chebyshev or Butterworth template function, the shapes of both IMN and OMN target gain functions are mathematically generated using NGF given in (19) as

$$T_{L-data-temp}(\omega) = T_{temp}(\omega), \ \omega_b \le \omega_i \le \omega_e \ , \quad (25)$$

$$T_{G-data-temp}(\omega) = T_{NGF-data}(\omega) \times T_{temp}(\omega) = \left(\frac{T_0}{|S_{21}(\omega)|^2} \right) \times T_{temp}(\omega), \ \omega_b \le \omega_i \le \omega_e \quad (26)$$

where T_{temp} denotes designer’s choice of bandpass Chebyshev or Butterworth template function as follows

$$T_{temp}(\omega) = \begin{cases} T_{ch}(\omega) \\ T_{Bw}(\omega) \end{cases}$$
$$\max\{T_{temp}(\omega)\} = 1, \quad \omega_b \le \omega \le \omega_e \ . \quad (27.i)$$

Bandpass Chebyshev or Butterworth template function $T_{temp}(\omega)$ is generated using bandpass to lowpass transformation $p \leftarrow (p^2 + \omega_0^2)/Bp = j(\omega^2 - \omega_0^2)/B\omega$ into lowpass prototype function $T_L(p)$ as

$$T_L(p) = \begin{cases} T_{ChL}(p) = \left. \dfrac{1}{1 + \varepsilon^2 T_m(x)^2} \right|_{x=p/j} \\ T_{BwL}(p) = \left. \dfrac{1}{1 + x^{2m}} \right|_{x=p/j} \end{cases}$$
$$T_{temp}(\omega) = T_L(p)\Big|_{p \leftarrow j\frac{\omega^2 - \omega_0^2}{\omega B}} \quad (27.ii)$$

where $T_L(p)$ is chosen by the designer as either a Chebyshev lowpass prototype function $T_{ChL}(p)$ built by *m*th order Chebyshev polynomials and ripple factor epsilon or a *m*th order Butterworth lowpass prototype function $T_{BwL}(p)$.

$T_{L-data-temp}(\omega) = T_{temp}(\omega)$ means the OMN target gain function shape will be a bandpass Chebyshev or Butterworth template function given with (27.i) [1].

$T_{G-data-temp}(\omega) = \left(T_0 / |S_{21}(\omega)|^2\right) T_{temp}(\omega)$ means the IMN target gain function shape will be a bandpass Chebyshev or Butterworth template function whose passband gain is a "tapered" shape determined by NGF gain $T_{NGF}(\omega) = \left(T_0 / |S_{21}(\omega)|^2\right)$ given in (19.b). In Fig. 2.a, typical gain shapes are shown.

Depending on the value of transistor S-parameters, $\alpha > 1$ situations may occur in (21) and $T_{L-data}(\omega)$ could be modeled by $T_{Ch}(\omega)$ or $T_{Bw}(\omega)$ function having maximum gain value greater than 1. However, for simplicity reasons, LRA assumes $\alpha = 1$ in our study. Designers should also pay attention to obtain sufficiently acceptable low values for IMN reflection function Γ_G along the operation band, while she/he tries to optimize to obtain least reflection values concurrently for OMN reflection function Γ_L.

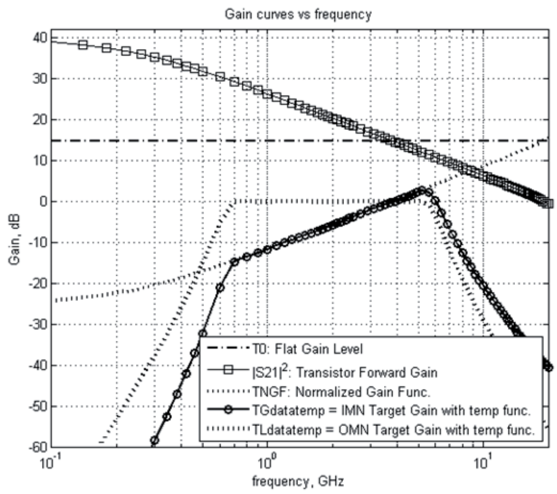
B. Balanced Reflection Approach (BRA): Although LRA almost always attains low reflection values (less than -10 dB) for OMN reflection function Γ_L along the operation band, it may not always yield a sufficiently low values for IMN reflection function Γ_G in all regions of the operation band. This may be caused by the fact that the S-parameter data of the transistor is not always suitable to obtain well-behaved, i.e. low valued, reflections in the whole operation band. BRA may help to share the OMN reflection into two parts almost equally between OMN and IMN, which may allow the input mismatch seen at IMN side to be healed. To achieve this, using (19.a), BRA assumes taking eth exponent of normalized gain function T_{NGF} to generate OMN and IMN target gain functions $T_{L-data}(\omega)$ and $T_{G-data}(\omega)$ using the following expressions

$$\begin{aligned} &T_{L-data}(\omega) = T_{NGF-data}(\omega)^e, \quad 0 \le e \le 1, \\ &T_{G-data}(\omega) = T_{NGF-data}(\omega)^{1-e}, \\ &T_{L-data}(\omega) = T_{G-data}(\omega) = \sqrt{T_{NGF-data}(\omega)}, \quad \text{if } e = 1/2 \end{aligned} \tag{27.iii}$$

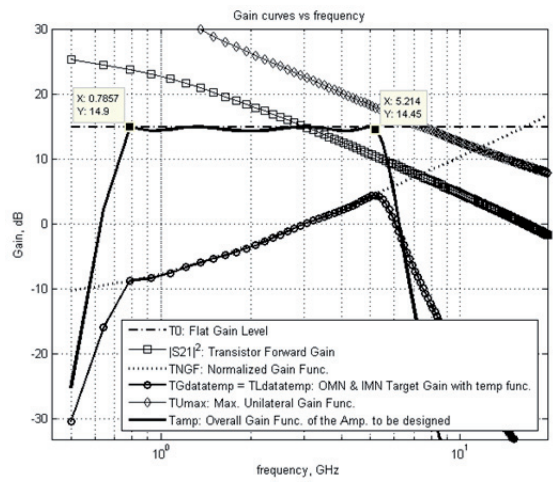
where the only situation that the *balanced* (or equal) reflections for both OMN and IMN can occur if $e = 1/2$ satisfied which corresponds to the geometric mean of the normalized gain function T_{NGF}. If we apply Chebyshev or Butterworth template modeling to (27.iii), we get the final forms of OMN and IMN target gain functions to be used in the optimization as

$$\begin{aligned} &T_{L-data-temp}(\omega) = T_{G-data-temp}(\omega) = \\ &= \sqrt{T_{NGF-data}(\omega)} \times T_{temp}(\omega) \\ &= \sqrt{\left(T_0 / |S_{21}(\omega)|^2\right)} \times T_{temp}(\omega), \ \omega_b \le \omega_i \le \omega_e \ . \end{aligned} \tag{27.iv}$$

In Fig. 2.b, typical gain shapes are shown for BRA approach.



a)



b)

Fig. 2. Typical gain curves for a) LRA and b) BRA.

4. A Brief Summary of SRFT

Simplified Real Frequency Technique (SRFT) solves the double matching problems by constructing lossless two-ports via optimization of the transducer power gain. In this process, lossless two-port is described in terms of its real normalized input reflection coefficient $\Gamma(p) = h(p)/g(p)$. The unknowns of the matching problem are selected as coefficients of $h(p)$ polynomial, i.e. $\{h_1, h_2, ..., h_n, h_{n+1}\}$. Then, "strictly Hurwitz" denominator polynomial $g(p)$ is uniquely determined from the initialized coefficients $\{h_1, h_2, ..., h_n, h_{n+1}\}$ by explicit factorization of the losslessness condition given by (9). In (9), polynomial $f(p)$ is specified by the user and it is constructed on the transmission zeros of the lossless matching network. Details of SRFT are omitted here. However, further and detailed information can be found in [7].

5. Wideband Amplifier Design Steps Using NGF-SRFT

NGF method combined with SRFT which might be called as NGF-SRFT, can be a good approach to design ultra wideband amplifiers. In a UWB microwave amplifier design, NGF method can be performed in 5 major steps: 1. OMN design, 2. IMN design, 3. synthesis, 4. reoptimization, 5. simulation.

This design method is given in the following in a step by step manner.

- Step 1: OMN design

1.1. The specifications of the amplifier to be designed are entered with parameters such as choice of template function either as Chebyshev or Butterworth i.e. $T_{temp}(\omega)=\{T_{ch}(\omega)\ or\ T_{Bw}(\omega)\}$, flat gain level $T_{data}(\omega)=T_0$, lower corner frequency f_L, upper corner frequency f_H, passband ripple ε, total number of L and C elements n for each of IMN and OMN, lower and upper optimization frequency bounds f_b and f_e.

1.2. Transistor forward gain is calculated by

$$T_{Q-data}(\omega)=|S_{21}(\omega)|^2, \quad \omega_b \le \omega \le \omega_e \tag{28}$$

using linear S-parameters of the transistor provided from the vendor in matrix form as

$$S(\omega)=\begin{bmatrix} S_{11}(\omega) & S_{12}(\omega) \\ S_{21}(\omega) & S_{22}(\omega) \end{bmatrix}, \quad \omega_b \le \omega \le \omega_e \tag{29}$$

Thus, its gain curve shape also becomes known, where ω_b and ω_e are lower and upper bounds of the optimization [1].

1.3. Eventually, using (19.b), the shape of normalized gain function T_{NGF} is computed by the ratio of given $T_{data}(\omega)=T_0$ and $|S_{21}(\omega)|^2$ as $T_{NGF}(\omega)=T_0/|S_{21}(\omega)|^2$. Then, the target gain functions to be used in the OMN and IMN optimization processes are generated using BRA as

$$\begin{aligned} T_{L-data-temp}(\omega) &= T_{G-data-temp}(\omega)= \\ &=\sqrt{T_{NGF-data}(\omega)\times T_{temp}(\omega)} \\ &=\sqrt{\left(T_0/|S_{21}(\omega)|^2\right)\times T_{temp}(\omega)},\ \omega_b \le \omega_i \le \omega_e \end{aligned}$$

1.4. OMN Optimization Code – OptOMN.m*:* This Matlab code deals with the computation of the unknown vectors $\boldsymbol{h_L}$, $\boldsymbol{f_L}$ and $\boldsymbol{g_L}$ of OMN. In the main code, before entering the OMN optimization code, the unknown vectors $\boldsymbol{h_L}$ and $\boldsymbol{f_L}$ are generally initialized in an ad-hoc manner, for example filled with unit values, i.e. $\boldsymbol{h_L}=[h_{L,1}\ h_{L,2}\ldots h_{L,n+1}]=\boldsymbol{I}=[1\ 1\ldots1\ 1]$ and $\boldsymbol{f_L}=[f_{L,1}\ f_{L,2}\ldots f_{L,n/2+1}\ldots f_{L,n+1}]=[0\ 0\ldots f_{L0}\ldots 0]=[0\ 0\ldots1\ldots 0]$. As will be detailed in the following, $\boldsymbol{g_L}$ need not to be initialized as "unknown" since it is totally derived in terms of unknown vectors $\boldsymbol{h_L}$ and $\boldsymbol{f_L}$ via losslessness condition of (9), (10). All coefficients of $\boldsymbol{f_L}$ vector are set to zero except the middle term $f_{L,n/2+1}=f_{L0}=1$, hence forming an even numerator polynomial $\boldsymbol{F_L}=\boldsymbol{f_L}\boldsymbol{f_L}^*=[F_{L,1}\ F_{L,2}\ldots F_{L,n/2+1}\ldots F_{L,n+1}]=[0\ 0\ldots F_{L0}\ldots 0]=[0\ 0\ldots f_{L0}^2\ldots 0]$. It is preferable to choose *ndc* as half of the element number of the OMN, i.e. $ndc=n/2$, to be able to obtain a symmetrical bandpass gain function for OMN with the form given as

$$T_L(p)=\frac{F_L(p)}{G_L(p)}=\frac{F_{L,0}p^{2ndc}}{G_{L,1}p^{2n}+G_{L,2}p^{2(n-1)}+\ldots+G_{L,n}p^2+G_{L,n+1}}. \tag{30}$$

Matlab optimization code OptOMN.m yields the resulting optimized polynomials h_L, f_L, g_L by executing the pseudo-code, given in Tab. 1, in step by step manner. n denotes the total element number of the bandpass OMN, *LHProots* denotes "Left-Half Plane roots" of strictly Hurwitz even

Optimization Code: OptOMN.m	*step*
$h_L=[h_1\ \ h_2\ldots h_n\ \ h_{n+1}],\quad h_L^*=h_L(-p)$ $f_L=[f_1\ \ f_2\ldots f_{n/2+1}\ldots f_n\ f_{n+1}]=[0\ 0\ldots f_{L,0}\ldots 0\ 0],$ $f_L^*=f_L(-p)=f_L(p)$	*i.*
$h_L=[h_1\ \ h_2\ldots h_n\ \ h_{n+1}],\quad h_L^*=h_L(-p)$ $f_L=[f_1\ \ f_2\ldots f_{n/2+1}\ldots f_n\ f_{n+1}]=[0\ 0\ldots f_{L,0}\ldots 0\ 0],$ $f_L^*=f_L(-p)=f_L(p)$	*ii.*
$G_L=H_L+F_L=[G_1\ \ G_2\ldots G_{n/2+1}\ldots G_n\ \ G_{n+1}]$ $=[H_1\ \ H_2\ldots(H_{n/2+1}+F_{n/2+1})\ldots H_n\ \ H_{n+1}]$	*iii.*
$g_L=[g_1\ \ g_2\ldots g_n\ \ g_{n+1}]=\sqrt{G_L(1)}poly(\ LH\mathrm{Pr}oots(G_L)\ )$	*iv.*
$\Gamma_L(p_i)=\dfrac{h_L(p_i)}{g_L(p_i)},\qquad p_i=j\omega_i,\ "i=1,2,\ldots,nd"$	*v.*
$T_L(p_i)=\dfrac{1-\lvert\Gamma_L(p_i)\rvert^2}{\lvert 1-S_{22}(i)\Gamma_L(p_i)\rvert^2},$	*vi.*
$d_L(i)=T_L(p_i)-T_{L-data-temp}(\omega_i)=T_L(i)-T_{temp}(\omega_i)$ $=T_L(p_i)-\left\{\sqrt{\left(T_0/\lvert S_{21}(\omega_i)\rvert^2\right)}\times T_{temp}(\omega_i)\right\}$ $\text{minimize}\left\{\sum_{i=1}^{nd}\lvert d_L(i)\rvert^2\right\}$ to yield h_L, g_L of $\Gamma_L=h_L/g_L$	*vii.*

Tab. 1. Pseudo code in OMN opt. func.

polynomial $G_L(p)$, *nd* represents the data number in the optimization range $[\omega_b\ldots\omega_e]$, $T_L(p_i)$ is the OMN gain function as given in (2), d_L is error or distance vector to be minimized and defined as the difference between the gain function $T_L(p_i)$ being optimized and the OMN target gain function $T_{L-data-temp}(\omega)$ being tracked. Matlab built-in function "conv" performs vector convolution or multiplication. In the Matlab optimization code development stage, among many optimization algorithm choices such as used in [27], [28], "fminsearch" [29] has been experimented and found as a highly successful and well-suited algorithm which takes the initial vector $\boldsymbol{x_{L0}}$= [$\boldsymbol{h_{L0}}$ $\boldsymbol{f_{L0}}$] as an input argument and yields always a convergent and realizable solution as $\boldsymbol{x_L}$= [$\boldsymbol{h_L}$ $\boldsymbol{f_L}$] optimized vector. In the main code, it is written as "x = fminsearch('OptOMN', x_{L0}, options);". At the end of the optimization, we have $h_L(p)$, $g_L(p)$ polynomials belonging to the input reflectance $\Gamma_L(p)=h_L(p)/g_L(p)$ of the OMN from which the normalized input impedance function $z_L(p)=a_L(p)/b_L(p)$ is obtained as given in (6) [1].

- Step 2: IMN design

2.1. IMN Optimization Code – OptIMN.m*:* The IMN optimization is very similar to that of OMN. Again, the optimization starts with the initial vectors $\boldsymbol{h_G}$=[$h_{G,1}$ $h_{G,2}$… $h_{G,n+1}$]=$\boldsymbol{I}$=[1 1…1 1] and $\boldsymbol{f_G}$=[$f_{G,1}$ $f_{G,2}$… $f_{G,n/2+1}$… $f_{G,n+1}$]=[0 0… f_{G0}… 0]=[0 0…1… 0]. Hence, F even numerator polynomial is formed as $\boldsymbol{F_G}$=$\boldsymbol{f_G f_G}$*=[$F_{G,1}$ $F_{G,2}$… $F_{G,n/2+1}$… $F_{G,n+1}$]= [0 0... F_{G0}... 0]=[0 0... f_{G0}^2... 0]. As in OMN, *ndc* is chosen as half of the total element number of IMN, i.e. *ndc* = *n*/2, to be able to obtain a symmetrical bandpass gain function for IMN with the form given as

$$T_G(p)=\frac{F_G(p)}{G_G(p)}=\frac{F_{G,0}p^{2ndc}}{G_{G,1}p^{2n}+G_{G,2}p^{2(n-1)}+...+G_{G,2n}p^2+G_{G,2n+1}}. \quad (31)$$

Matlab optimization code OptIMN.m yields the resulting optimized polynomials h_G, f_G, g_G by executing the pseudo-code, given in Tab. 2, in a step by step manner. $T_{G\text{-}data-temp}(\omega)$ is the tapered gain form as seen in Fig. 1.b. In the main code, it is written as "x=fminsearch('OptIMN', x_{G0}, options);". At the end of the optimization, we have $h_G(p)$, $g_G(p)$ polynomials belonging to the input reflectance $\Gamma_G(p)=h_G(p)/g_G(p)$ of the IMN from which the normalized input impedance function $z_G(p)=a_G(p)/b_G(p)$ is obtained as given in (6) [1].

- Step 3: Synthesis

Any rational PR (positive real) driving point input impedance function $z(p)$ of the form (6) can always be realized as LC lossless ladder network with resistive termination in Darlington's sense. In principle, an LC ladder synthesis is achieved by means of a straightforward long division process (see (7) and (8)) of an immitance (impedance or admittance) function. At each step, a pole at DC or infinity is removed. After each step, the degree of the remaining function is reduced. This process continues until we end up with a constant term [26], which determines the termination resistance of the corresponding matching network. When the gain function form given in (18) is considered, for each of IMN and OMN matching networks, integer n designates the total number of reactive elements when $z(p)$ is synthesized as a lossless 2-port in resistive termination. *ndc* is the number of DC transmission zeros which are realized as series capacitors and shunt inductors in a 2-port. $n_\infty=n-n_{dc}>0$ is the number of transmission zeros at infinity which are realized as series inductors and shunt capacitors. In [26], much more detailed information can be found about the "high precision synthesis of bandpass LC ladders" together with very useful Matlab synthesis package.

Choosing *ndc* = *n*/2 allows one to be able to create "symmetrical" roll-off characters for gain functions $T_G(p)$ and $T_L(p)$ of IMN and OMN, respectively. For the design example worked in Section 6, at the beginning of the design, element number for each matching network is chosen, for example, as $n=8$ excluding termination resistances denoted by R9. As seen in Fig. 5 of this design example, we synthesize both OMN and IMN input impedance functions in such a way that each is yielded as bandpass LC lossless ladder having *ndc* = $n/2=4$ number of hp (high-pass) sections with series capacitors and shunt inductors (C1, L2, C3, L4) and $n_\infty=n-n_{dc}=n/2=4$ number of lp (lowpass) sections with series inductors and shunt capacitors (C8, L7, C6, L5).

Optimization Code: OptIMN.m	*step*
$h_G=[h_1\ \ h_2...h_n\ \ h_{n+1}],\quad h_G^*=h_G(-p)$ $f_G=[f_1\ \ f_2...f_{n/2+1}...f_n\ f_{n+1}]=[0\ 0...f_{G,0}...0\ 0]$ $f_G^*=f_G(-p)=f_G(p)$	*i.*
$H_G=conv(h_G,h_G^*)=\left[H_1\ H_2...H_n\ H_{n+1}\right]$ $F_G=conv(f_G,f_G^*)=\left[F_1\ F_2...F_{n/2+1}...F_n\ F_{n+1}\right]$ $=\left[0\ 0...f_{G,0}^{\ 2}...0\ 0\right]$	*ii.*
$G_G=H_G+F_G=[G_1\ \ G_2...G_{n/2+1}...G_n\ G_{n+1}]$ $=[H_1\ \ H_2...(H_{n/2+1}+F_{n/2+1})...H_n\ H_{n+1}]$	*iii.*
$g_G=[g_1\ \ g_2...g_n\ g_{n+1}]=\sqrt{G_G(1)}poly(\ LH\,\mathrm{P}roots(G_G)\ )$	*iv.*
$\Gamma_G(p_i)=\dfrac{h_G(p_i)}{g_G(p_i)},\qquad p_i=j\omega_i,\ "i=1,2,...,nd"$	*v.*
$\Gamma_{IN}(i)=S_{11}(i)+\dfrac{S_{12}(i)S_{21}(i)\Gamma_L(p_i)}{1-S_{22}(i)\Gamma_L(p_i)}$ $T_G(i)=\dfrac{1-\lvert\Gamma_G(p_i)\rvert^2}{\lvert 1-\Gamma_{IN}(i)\Gamma_G(p_i)\rvert^2}$	*vi.*
$d_G(i)=T_G(p_i)\text{ - }T_{G-data-temp}(\omega_i)$ $=T_G(p_i)\text{ - }\left\{\sqrt{\left(T_0/\lvert S_{21}(\omega_i)\rvert^2\right)}\times T_{temp}(\omega_i)\right\}$ minimize$\left\{\sum_{\mathrm{i=1}}^{\mathrm{nd}}\lvert d_G(i)\rvert^2\right\}$ to yield h_G,g_G of $\Gamma_G=h_G/g_G$	*vii.*

Tab. 2. Pseudo codes in IMN opt. func.

In the Matlab code "Main_NGF_SRFT.m", a function [CVal, CType]=general_synthesis(a, b) is executed twice to synthesize each of the normalized input impedance functions $z_L(p)=a_L(p)/b_L(p)$ and $z_G(p)=a_G(p)/b_G(p)$, respectively. The function "general_synthesis" performs the synthesis procedure based on the theoretical description introduced at the beginning of this step (1st paragraph), i.e. Step 4. Once the OMN optimization is completed, the optimized coefficient vectors $\mathbf{a}_L$ and $\mathbf{b}_L$ are loaded into the function [CVal_L, CType_L]=general_synthesis(aL, bL), which synthesizes the impedance function $z_L(p)=a_L(p)/b_L(p)$ and eventually yields normalized component values vector as CVal_L=[1.2701 9.0262 2.3485 18.0363 4.1510 0.4823 2.2512 0.0871 2.6377] and topology vector in terms of component types as CType_L=[2 7 2 7 1 8 1 8 9]; where 2 and 7 denotes series-C and shunt-L; 1 and 8 denotes series-L and shunt-C; 9 denotes termination resistance. Similarly, after the IMN optimization is completed, re-running the function as [CVal_G, CType_G]=general_synthesis(aG, bG) results the normalized component values of the IMN network and its topology such that CVal_G=[2.9015 5.6189 3.9557 9.6290 1.7046 0.5265 0.2497 0.1894 1.5548] and CType_G=[2 7 2 7 1 8 1 8 9], respectively. Once the normalized element values for each matching network are obtained by the synthesis process; then, the actual element values for OMN and IMN are computed, as seen in Tab. 3.a, via a denormalization process that uses impedance normalization factor with R0 = 50 Ohms and frequency normalization factor with $f_{norm} = 2\pi f_H = 2\pi$ (5.2 GHz).

- Step 4: Reoptimization

Reoptimization of the designed amplifier can be done to be able to obtain a new amplifier having 50 Ohm resistive terminations for both OMN and IMN networks. One must observe that the performance of the reoptimized amplifier does not go far away from the desired technical specs prescribed at the beginning of the design. This can be a validation condition; if it is not possible to meet this condition, a 50 Ohm reoptimized design may not be preferred over non-50 Ohm design. However, 50-Ohm termination, if it could be achieved via reoptimization, is a desirable situation for many practical problems; moreover, it eases the measurement of the amplifier since its input and output are compatible to the standard 50 Ohm ports of the measurement equipment such as VNAs (vector network analyzers).

We should mention that, when the analytic form of the normalized immitance (impedance $z(p)$ or admittance $y(p)$) is synthesized as a lossless two-port in resistive termination, one may end up with a transformer depending on the value of the termination resistance. 'Transformerless' or 50 Ohm termination is always desired in the final synthesis, however, this can only be achieved for low-pass design problems (see p. 361 [7]). Indeed, at DC, i.e. when $p = j\omega = 0$, $z_M(p) = (1+\Gamma_M(p))/(1-\Gamma_M(p)) = (g_M(p)+h_M(p))/(g_M(p)-h_M(p))$ becomes $z_M(p) = (g_{n+1,M}+h_{n+1,M})/(g_{n+1,M}-h_{n+1,M})$ and, by substituting $h_{n+1,M} = 0$, it results $z_M(p) = g_{n+1,M}/g_{n+1,M} = 1$. This means that, for low-pass designs only, each of the IMN and OMN normalized input impedances has 1 Ohm termination resistance that corresponds to standard 50 Ohm ("transformerless" design).

To our knowledge, for bandpass design problems, it is a challenging task to obtain 50 Ohm terminations via an analytic procedure similar to the above mentioned method for lowpass designs. On the other hand, reoptimization would be impossible if the topology of the designed amplifier was not available. This is owed to the SRFT that determines the topology of the non-50 Ohm amplifier as the initial design to be able to pass to the reoptimization stage if desired.

Once the topology of the initial non-50 Ohm amplifier (seen in Fig. 5) is obtained via SRFT technique by the Matlab code "Main_NGF_SRFT.m", then, to be able to obtain a 50 Ohm terminated amplifier in MWO (of AWR Corp.) environment, both termination resistances (two R9s in Fig. 5) are firstly kept constant as 50 Ohm. Then, all LC elements of IMN and OMN are reoptimized in such a way that the overall gain function is tracked as much precise as possible. As a result, Tab. 3.a shows the element values of the non-50 Ohm amplifier, whereas Tab. 3.b shows the element values of the reoptimized, in other words 50-Ohm terminated, amplifier. In Fig. 4, a high agreement between the gain performances of both non-50 Ohm and the reoptimized amplifier can be seen.

- Step 5: Simulation

The amplifier composed of transistor Q and designed matching networks IMN/OMN is simulated in the MWO (Microwave Office, AWR Corp.) environment [23]. Theoretical and simulated gain performances are compared to observe the degree of agreement between them (see Fig. 3 and Fig. 4 to compare the performances of the theoretical Matlab design with that of the simulation in Fig. 4 (dashed one)).

6. Design of an UWB Amplifier and Its Performance Evaluations

Design example: UWB amplifier based on NGF-SRFT

A flat gain UWB amplifier is to be designed with the following specifications: flat gain level $T_{data}(\omega) = T_0 = 14.90\ dB$ (to assure the stability, T_0 is selected inside the lower region below the $T_{U\max}$ maximum unilateral power gain curve which is a perfect match case under the assumption: $\Gamma_L = S_{22}^*$, $\Gamma_G = S_{11}^*$ if $S_{12} = 0$), lower corner frequency $f_L = 0.8$ GHz, upper corner frequency $f_H = 5.2$ GHz, Chebyshev template function $T_{temp}(\omega) = T_{ch}(\omega)$, passband ripple factor $\varepsilon = 0.27$ (0.3056 dB), total number of L and C elements $n = 8$ (excluding termination resistance) for each of IMN and OMN, $ndc = n/2 = 4$ transmission zeros at DC, lower and upper optimization frequency bounds $f_b = 0.5$ GHz and

$f_e = 26$ GHz, active device TGF2021-01 of TriQuint Semiconductor, 1.4 W wideband DC-12 GHz discrete power pHEMT [22].

Note that, choosing the total number of L and C elements as $n = 8$, which is a relatively high filter order, can make possible to obtain IMN and OMN networks belonging to such an amplifier that its gain characteristic can resemble much more closely the shape of the given prescribed target gain function along a wide frequency band

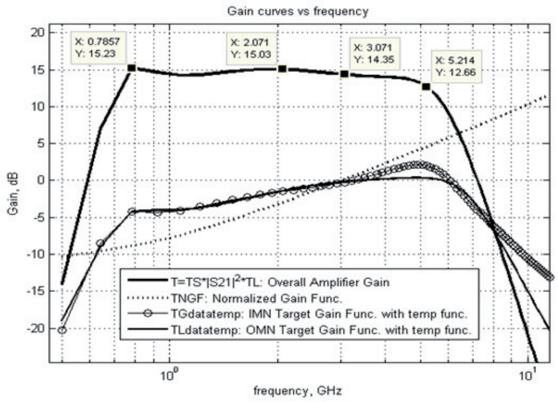


Fig. 3. Matlab design: shapes of typical gain functions for the designed UWB amplifier.

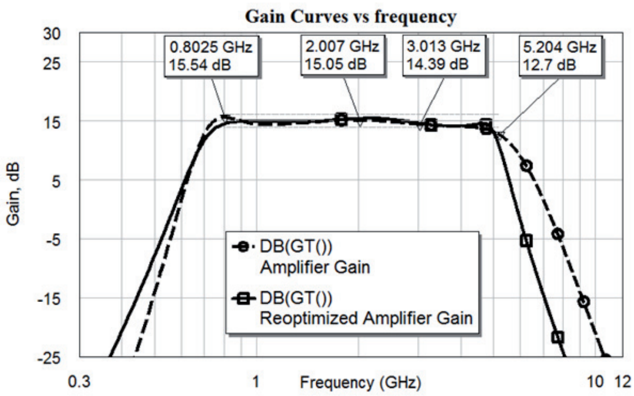


Fig. 4. MWO simulation: gain performances of the designed UWB amplifier by Matlab code (dashed) and reoptimized one by MWO (solid).

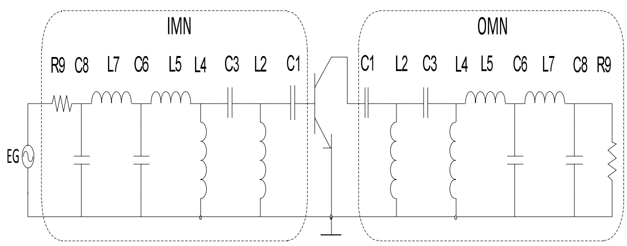


Fig. 5. Matlab design: schematic of the designed UWB amplifier.

of operation, (0.5-26) GHz. According to our experience with the Matlab code "Main_NGF_SRFT.m", element number n smaller than $n = 8$, could not yield matching networks that can resemble thoroughly the desired prescribed amplifier gain shape over a wide frequency band of operation. On the other hand, we have observed that, for the higher element number n greater than 12, severe accumulated numerical errors in the computations yield distorted shapes for the target gain functions $T_{G\text{-}data\text{-}temp}(p)$ and $T_{L\text{-}data\text{-}tem}(p)$, which cannot be used as the optimization target functions at all. The main idea in choosing the "suitable" filter order n is that; not only this order n should yield matching networks satisfying the desired gain shape precisely as much as possible along a wide frequency band, but also it concurrently should result such matching circuits with minimal number of elements as much as possible.

Element Values: pF, nH, Ohm	Matching Network		Element Values: pF, nH, Ohm	Matching Network	
	IMN	OMN		IMN	OMN
C1	1.759	0.770	L5	2.584	6.292
L2	8.517	13.682	C6	0.319	0.292
C3	2.398	1.424	L7	0.379	3.412
L4	14.595	27.339	C8	0.115	0.053
			R9	**77.739**	**131.884**

a)

Element Values: pF, nH, Ohm	Matching Network		Element Values: pF, nH, Ohm	Matching Network	
	IMN	OMN		IMN	OMN
C1	1.739	1.200	L5	2.464	8.542
L2	6.417	15.08	C6	0.5745	0.3404
C3	3.898	1.404	L7	0.2435	5.812
L4	198.0	614.7	C8	0.1228	0.05208
			R9	**50.00**	**50.00**

b)

Tab. 3. a) Matlab design: element values of the designed UWB amplifier by Matlab, b) element values of the "reoptimized" amplifier by MWO (of AWR Corp.).

Lower and upper optimization frequency bounds are set as $f_b = 0.5$ GHz and $f_e = 26$ GHz, between which scattering parameters measured at 52 different frequencies are given in the S-parameter file of the active device TGF2021-01. By using interpolation, the optimization band in (0.5-26) GHz is further divided into $nd = 714$ equal frequency steps, each measures $\Delta f = 35.714$ MHz. Therefore, in every stage of the design, the computations are done with frequency increments of $\Delta f = 35.714$ MHz, enabling high resolution computations.

Solution: Using the developed Matlab code "Main_NGF_SRFT.m", starting with initial polynomials h_L, f_L, h_G, f_G, OptOMN.m and OptIMN.m optimization codes yield the following optimized results:

$$h_L(p) = 0.7986p^8 + 3.1527p^7 + 4.1501p^6 + 4.4580p^5 + 2.6134p^4 + 1.3310p^3 + 0.3189p^2 + 0.0335p + 0.0042$$
$$f_L(p) = 2.5065p^4$$
$$g_L(p) = 0.7986p^8 + 3.8030p^7 + 6.9821p^6 + 8.5300p^5 + 5.8964p^4 + 2.1443p^3 + 0.4175p^2 + 0.0442p + 0.0042$$

$$h_G(p) = 0.0598p^8 + 0.1512p^7 + 1.6385p^6 + 0.3535p^5 + 0.8212p^4 + 0.3114p^3 + 0.1434p^2 + 0.0103p + 0.0023$$
$$f_G(p) = 2.2607p^4$$
$$g_G(p) = 0.0598p^8 + 0.2551p^7 + 1.9914p^6 + 3.3822p^5 + 3.6454p^4 + 1.2133p^3 + 0.2415p^2 + 0.0235p + 0.0023$$

then, numerator and denominator polynomials of Darlington input impedance functions $z_L(p) = a_L(p)/b_L(p)$ and $z_G(p) = a_G(p)/b_G(p)$ for the corresponding OMN and IMN networks are

$$a_L(p) = 1.5972p^8 + 6.9557p^7 + 11.1322p^6 + 12.9881p^5 + 8.5098p^4 + 3.4753p^3 + 0.7364p^2 + 0.0777p + 0.0084$$
$$b_L(p) = 0.6503p^7 + 2.8320p^6 + 4.0720p^5 + 3.2830p^4 + 0.8133p^3 + 0.0986p^2 + 0.0106p$$

$$a_G(p) = 0.1196p^8 + 0.4063p^7 + 3.6299p^6 + 3.7357p^5 + 4.4667p^4 + 1.5247p^3 + 0.3848p^2 + 0.0338p + 0.0045$$
$$b_G(p) = 0.1039p^7 + 0.3529p^6 + 3.0288p^5 + 2.8242p^4 + 0.9019p^3 + 0.0981p^2 + 0.0132p$$

And finally, optimized partial gain functions in p-domain for OMN and IMN networks in the form of (30) and (31) are

$$T_L(p) = \frac{F_L(p)}{G_L(p)} = \frac{F_{L,0}p^{2ndc}}{G_{L,1}p^{2n} + G_{L,2}p^{2(n-1)} + \ldots + G_{L,n}p^2 + G_{L,2n+1}} = \frac{6.2825p^8}{G_L(p)}$$
$$G_L(p) = 0.6378p^{16} - 3.3109p^{14} - 6.7120p^{12} - 6.0658p^{10} + 3.6870p^8 - 0.3688p^6 + 0.0344p^4 + 0.0016p^2 + 1.7570\times10^{-5}$$

$$T_G(p) = \frac{F_G(p)}{G_G(p)} = \frac{F_{G,0}p^{2ndc}}{G_{G,1}p^{2n} + G_{G,2}p^{2(n-1)} + \ldots + G_{G,n}p^2 + G_{G,2n+1}} = \frac{5.1105p^8}{G_G(p)}$$
$$G_G(p) = 0.0036p^{16} + 0.1732p^{14} + 2.6761p^{12} + 2.4893p^{10} + 6.0318p^8 + 0.1386p^6 + 0.0179p^4 + 0.0005p^2 + 5.1514\times10^{-5}$$

which construct the overall amplifier gain function as $T(\omega) = T_G(\omega)\left|S_{21}(\omega)\right|^2 T_L(\omega)$ by substituting $p = j\omega$. Typical gain functions for the designed UWB amplifier via Matlab code Main_NGF_SRFT.m, the amplifier schematic and the element values are given in Fig. 3, Fig. 5 and Tab. 3.a, respectively. For the NA (Network Analyzer) measurement requirements, R9 termination resistances of both IMN and OMN should be replaced by transformers whose winding ratios are determined in such a way that the generator and the load each sees 50 Ohm standard impedance of NA ports. Although termination resistances were optimized to values different than 50 Ohm, reoptimization for both OMN and IMN can yield 50 Ohm terminations hence the need for transformers is avoided.

As recalled from the step 4, in Section 5, the reoptimization issue was first mentioned and elaborated extensively. If we re-state, Tab. 3.a and Tab. 3.b show the element values of the non-50 Ohm amplifier designed by the Matlab code "Main_NGF_SRFT.m" and the reoptimized amplifier by the MWO, respectively. In Fig. 4, a high agreement between the gain performances of both non-50 Ohm and the reoptimized, i.e. 50-Ohm terminated, amplifier is shown.

The amplifier gain yielded by the simulation in MWO is seen in Fig. 4 (dashed one) and it is in %100 agreement with that of theoretical gain obtained via Matlab code, as seen in Fig. 3. The overall amplifier gain flatness is acceptable over the operation band and its flatness may be improved depending on the desired degree of the reflections at OMN and IMN as far as the S-parameter data of the transistor permits. For those who desire to reproduce the results of the design example via Matlab code Main_NGF_SRFT.m, we recommend them to run the codes given in [30].

6.1 Stability Considerations

Absolute and conditional stability checking of the designed amplifier can be done using the following equation set

$$R_{OUT}(\omega) \geq 0,\ \forall\omega \qquad R_{IN}(\omega) \geq 0,\ \forall\omega \tag{32.i}$$

$$R_{OUT}(\omega) + R_L(\omega) \geq 0,\ \forall\omega \qquad R_{IN}(\omega) + R_G(\omega) \geq 0,\ \forall\omega \tag{32.ii}$$

where R_{OUT}, R_{IN}, R_L, R_G are resistive (or real) parts of the impedances seen at transistor output, transistor input, OMN input, IMN output, respectively. (32.i) is used for absolute stability checking and (32.ii) for conditional stability checking of the amplifier [7]. Above equation set is computed along the whole band, i.e. [0.5-26] GHz given for the transistor TGF2021-01, and the results are seen in Fig. 6. As seen from the figure, all the computed normalized resistive parts of the impedances are always greater than zero along the design band [0.8-5.2] GHz, that is to say they have not any negative resistances. Therefore, we can say that the amplifier is absolutely stable.

6.2 Computation of h and g Polynomial Coefficients via Error Minimization

Computation of the coefficients of *h* and *g* polynomials for both OMN and IMN is the main aim of each of the optimization functions OptOMN.m and OptIMN.m. As seen in *Step vii* of each of Tab. 1 and Tab. 2, at any iteration instant *t*, the optimization function computes the gain function $T_M(p_i)$ (M={G or L} and $p_i = j\omega_i = j2\pi f_i$) as a vector in $1\times nd$ dimension, where $i = 1..nd$. For the example design in Section 6, *nd* is taken as 714. The computation of $\boldsymbol{T_M}(p_i)$ vector is done within a loop in which the frequency variable f_i is swept in the frequency range $f_b \le f_i \le f_e$. As stated earlier, the optimization band is given as $(f_b - f_e) = (0.5\text{-}26)$ GHz composed of $nd = 714$ number of total frequency increments, each measures $\Delta f = 35.714$ MHz. Then, the computed gain vector $\mathbf{T_M}$ is substracted from the same-length (i.e. *1xnd*) target gain function vector $T_{M-data-temp}(\omega_i)$ to form the error vector $\boldsymbol{d_M}$ in $1 \times nd$ dimension. Therefore, at any iteration instant *t*, for every point f_i in the frequency axis of length *nd*, a corresponding number $d_M(i)$ that measures the distance

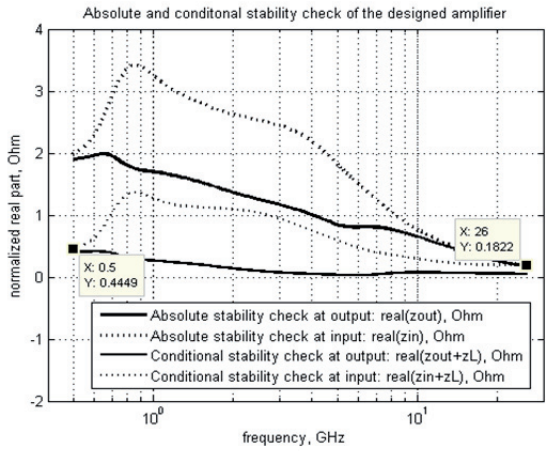


Fig. 6. Stability check of the designed amplifier.

between the computed mathematical gain and the target gain curve is created. In *step vii* of either Tab. 1 or Tab. 2, the error minimization stated by a formula as the "summation of the squares of the errors (SSE)" acts a very critical role in that the nonlinear optimization algorithm (NOA) could converge to a successful solution, i.e. finding the optimized solution for $\mathbf{h}_M$ and $\mathbf{g}_M$ vectors, each in $1\times(n+1)$ dimension. For any iteration instant *t* in the related optimization function (in OptOMN or OptIMN), a scalar number $SSE(t)$ can be computed in accordance with "summation of the square of the error vector elements" along the optimization frequency band such that

$$SSE(t) = \sum_{i=1}^{nd} \left| d_M(i) \right|^2, \quad i = 1..nd \tag{33}$$

$$\text{where } d_M(i) = \begin{cases} d_L(i) = T_L(p_i) - T_{L-data-temp}(\omega_i), \text{if } M = L \\ d_G(i) = T_G(p_i) - T_{G-data-temp}(\omega_i), \text{if } M = G \end{cases}$$

This scalar is a number that uniquely defines the measure of the closeness or the similarity degree of the shape of the current gain computed at the t^{th} instant to the target gain curve being tracked. The smaller the SSE scalar, the closer the computed gain shape to the target gain shape. fminsearch nonlinear optimization algorithm ends the optimization upon the current $SSE(t)$ nears a sufficiently small SSE_{val} (which is set to $TolX = 10^{-4}$ in the "options" of the fminsearch algorithm) and returns to the main code after loading the optimized $\boldsymbol{h}$ and $\boldsymbol{f}$ vectors into the solution vector as $\vec{x} \leftarrow [\vec{h}_{\text{opt}} \ \ \vec{f}_{\text{opt}}]$. Once the main code reloads the solved $\boldsymbol{h}$ and $\boldsymbol{f}$ vectors as $\vec{h} \leftarrow x(\vec{h}_{\text{opt}}),\ f \leftarrow x(\vec{f}_{\text{opt}})$, $\boldsymbol{g}$ vector is computed using *step iii* of Tab. 2 (or Tab. 1). As a result, final $\boldsymbol{h}$ and $\boldsymbol{g}$ solution vectors have been computed. To be able to comprehend thoroughly how "minimization of the summation of the square of the errors", i.e. the scalar *SSE*, acts an effective role in the success of the optimization algorithm convergence that eventually yields solved $\boldsymbol{h}$, $\boldsymbol{f}$ and $\boldsymbol{g}$ vectors; we consider the IMN optimization case of the design example in Section 6. For this design example, in the IMN optimization function OptIMN.m, the computed scalar numbers $SSE(t)$ with respect to the iteration number *t*, are shown in Fig. 7. As seen in the figure, the IMN optimization has ended at the iteration instant t_{6514}. Thus, the final solution belongs to the iteration instant when *t* is equal to 6514. Therefore, the solution vector becomes $\vec{x} \leftarrow [\vec{h}_{\text{opt}} \ \ \vec{f}_{\text{opt}}] = [\vec{h}_{6514} \ \ \vec{f}_{6514}]$. As noticed, the solution has the least SSE number which is $SSE_{6514} = 0.1004$. For the marked instants at $t = \{1, 3570, 4500, 5400, 6300, 6514\}$ in Fig. 7, the $\boldsymbol{h}_{\text{G}}$ and $\boldsymbol{g}_{\text{G}}$ vectors

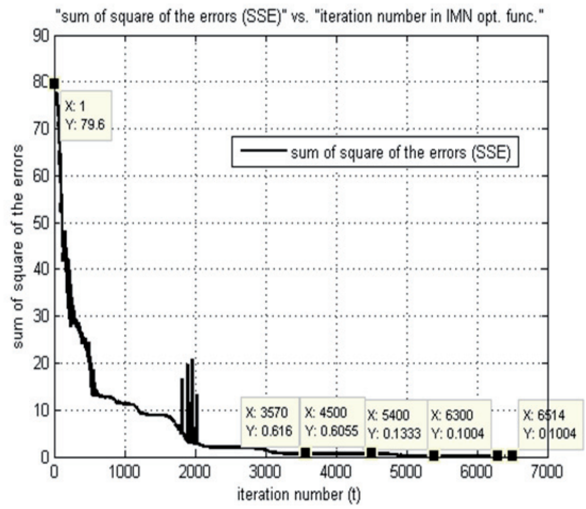


Fig. 7. Deviation of the "sum of the square of the errors (*SSE*)" with respect to the iteration number (*t*) in the IMN optimization.

Iter. Num. (t)	*SSE (t)*	$h_G(t)=[h_1\ h_2\ h_3\ h_4\ h_5\ h_6\ h_7\ h_8\ h_9]$
1	79.6	1 1 1 1 1 1 1 1 1
3570	0.616	0.8043 0.8331 2.8350 0.5265 1.5651 0.3186 0.2192 0.0113 0.0040
4500	0.6055	0.7838 0.8413 2.8075 0.5174 1.5865 0.3020 0.2310 0.0108 0.0043
5400	0.1333	0.1555 0.5371 1.8235 0.5379 0.9188 0.3544 0.1504 0.0111 0.0024
6300	0.1004	0.0597 0.1508 1.6381 0.3533 0.8210 0.3113 0.1433 0.0103 0.0023
6514	0.1004	0.0598 0.1512 1.6385 0.3535 0.8212 0.3114 0.1434 0.0103 0.0023

Tab. 4. Solved **h** vector with respect to iteration number.

Iter. Num. (t)	*SSE (t)*	$g_G(t)=[g_1\ g_2\ g_3\ g_4\ g_5\ g_6\ g_7\ g_8\ g_9]$
1	79.6	1.0000 4.8492 12.2573 20.3230 23.9289 20.3230 12.2573 4.8492 1.0000
3570	0.616	0.8043 2.1437 5.2603 5.9865 4.7699 1.5089 0.3377 0.0327 0.0040
4500	0.6055	0.7838 2.1214 5.2268 5.9710 4.7926 1.5325 0.3509 0.0339 0.0043
5400	0.1333	0.1555 0.8052 2.9805 4.4202 4.0774 1.3155 0.2555 0.0249 0.0024
6300	0.1004	0.0597 0.2546 1.9902 3.3810 3.6448 1.2131 0.2414 0.0235 0.0023
6514	0.1004	0.0598 0.2551 1.9913 3.3821 3.6454 1.2133 0.2415 0.0235 0.0023

Tab. 5. Solved **g** vector with respect to iteration number.

are computed using the solution vector $\boldsymbol{x}$ and are shown in Tab. 4, Tab. 5, respectively. Notice how close all the values of the coefficients of $\boldsymbol{h}$ and $\boldsymbol{g}$ vectors having SSE values sufficiently close to the final SSE_{6514} where the optimization ends with the solution.

7. Conclusion

In this work, a UWB amplifier design methodology based on Simplified Real Frequency Technique (SRFT), we name as NGF-SRFT, is proposed and used in the design of an ultra wideband (800-5200 MHz) low/medium power microwave amplifier that can cover all frequency bands of the communication standards. NGF method gives the designer a substantial approach to compute the target gain functions precisely to be tracked by the OMN and IMN optimization processes. In other words, the fineness of NGF comes from its capability of sharing of overall amplifier gain function into two separate mathematically generated target gain functions in a simple way but relied on a mathematical base. Therefore, in NGF-SRFT, while NGF provides the generation of two target gain functions required by the SRFT assisted nonlinear optimization processes; SRFT, which has essentially "topology-free" nature, allows numerical computation of input impedances of IMN/OMN as realizable and synthesizable positive real functions. NGF-SRFT not only permits to design the matching networks of the amplifier in ladder topology composed of L and C elements, but also helps the resulting amplifier to track the desired overall gain function shape, as seen in Fig. 3, provided that the maximum power transfer is attained as much and concurrent as possible.

The theoretical results generated by the Matlab code are found in 100% agreement with the results obtained from the simulations done in MWO environment. The practical implementation of (0.8-5.2) GHz UWB amplifier based on NGF-SRFT methodology is currently under investigation. If one stays loyal to the fundamentals of lumped element design given in the paper, the NGF-SRFT method does not differ much to design all-distributed element based design using various technologies such as microstrip technology. All-distributed element based design using Richard transformations [7] via NGF-SRFT is currently worked both in theoretical and practical aspects.

Acknowledgements

We thank to Ahmet Aksen, Professor and Head of the Department of Electronics Engineering of Işık University, for his valuable contributions and supportive comments.

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